Underwater Acoustic Modem Using Multi-Carrier Modulation

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Abstract - An underwater digital acoustic communications system based on multicarrier modulation technique is presented. This technique is relatively robust in a strong multipath fading environment and has been successfully tested in a shallow water channel at a data rate up to 10kbps over 1km. The test system using 48 carrier frequencies for transmitting 48 parallel bits of data in each packet. The data stream is encoded into a sequence of packets. These packets are encapsulated by headers, synchronisation and training signals. In order to reduce the multipath fading problem, an adaptive channel equalisation scheme with a LSM algorithm is used. The multipath fading problem can be further improved by a redundant carrier assignment and wavelet based MFM scheme. The advantage of this scheme is that the fading is reduced without affecting the channel capacity.

I. INTRODUCTION

There is a growing demand for underwater wireless communication with the rapid expansion in marine environmental monitoring activities. Much underwater instrumentation requires high speed data rates over a relatively long distance, in terms of kilometres, in a shallow water environment. The most common medium of underwater wireless communication is ultrasonic waves. With communication in shallow water environments, factors such as multipaths and shipping noise have to be taken into account. Thus, the main problem with using ultrasonic communication underwater is the complexity of the water channel. Channel imperfections are numerous and, as well as multipaths and water motion, can be due to density gradients, and the non-homogeneity of the water due to particles of solid or gaseous matter.

Multipaths can exist in almost any transmission medium, whether bounded or unbounded. Reflections from boundaries can create signal paths between transmitter and receiver in addition to the direct path. In unbounded media, it is feasible that spurious signal paths may be created by temperature or velocity gradients. The effect of temperature gradients can be to refract a wave front such that the signal is bent round to the point of reception, so interfering with the direct signal. Similarly, a velocity gradient lying across the axis of wave propagation can cause what appears to be refraction, bending the wave front. In a shallow, open water environment, the multipath problem is mainly due to reflections from the surface of the water and the sea or lake floor. Previous work at City University of Hong Kong [1-4] has shown that it is possible to send data reliably through liquid filled pipes, where the multipath problem is considerable.

Underwater communication using ultrasound has progressively improved, both in terms of data rate and distance. Due to the harsh environment often encountered, many well known techniques cannot be applied and therefore it is a compromise between speed, reliability and distance [5]. Table 1 shows the results obtained by different methods based on current experimental results. It can be seen that the distance and data rate for a water depth of 10m and less are lower than for deeper water. The modem reported in this paper is developed for marine parks in Hong Kong, where water depth averages around 10m, and within 2km of the shoreline. These are the design parameters for the system described in this paper, with a data rate of 10kbps so that compressed audio and video data can be transmitted, along with colour and monochrome still pictures. This paper only concentrates on the transport layer of this modem. The error correction, data compression are considered as the outer layers, and are not reported here.

TABLE 1
REPORTED RESULTS OF UNDERWATER ACOUSTIC COMMUNICATION

Developed by	Water depth (m)	Carrier frequency (Hz)	Distance (km)	Modulation	Data rate (bps)
M. Stovanovic et al [6]	100-200	25k	3	QPSK	10k
L. E. Freitag et al [7]	6-18	25k	0.7	MFSK	5k
J. A. Catipovic et al [8]	6-20	20k	0.75	128-FSK	10k
H. K. Yeo et al [9]	~ 18	10k	5 (max)	QPSK / BPSK	4k
H. A. Leinhos [10]	~ 13	3.5k	~ 6	1870-coded QPSK	1250 symbols per second

II. MULTICARRIER MODULATION

Multicarrier modulation divides a channel into a set of parallel independent subchannels [12]. The SNR of each subchannel is measured and a suitable number of bits is then assigned to each channel. There are two reasons for choosing Multicarrier Modulation (MCM) in the system. According to Bingham [13], the MCM signal can be processed in a receiver without the enhancement of noise or interference that is caused by linear equalisation of a single-carrier signal. Another is that the long symbol time used in MCM produces a much greater immunity to impulse noise and fast fades. Bingham [13] and Proakis [14] show that if the input data is Mf_s b/s. and they are grouped into blocks of M bits at block (symbol) rate of f_s,

$$f_{c,n} = n\Delta f \text{ for } n = n_1 \text{ to } n_2 \tag{1}$$

$$M = \sum_{n=n_1}^{n_2} m_n \tag{2}$$

where $N_c = n_2 - n_1 + 1$, $f_{c,n}$ = carrier frequency, Δf = frequency separation and N_c = number of carriers.

In the system described here, the modulation and demodulation techniques used are IFFT and FFT. IFFT and FFT are well-known efficient algorithms and significantly reduce the complexity of implementing the modulation and demodulation functions. The benefit in system implementation using IFFT and FFT are mentioned by Lee et al. [12] and Rizos et al.[15]. However, the resulting signalling filters have relatively large overlapping sidelobes (-13dB), and this causes a deviation from the ideal multicarrier scheme of independent carriers.

In this system, the binary input data are parsed to each subchannel by encoding them into a series of fixed number of bits symbols. The number of bits for each subchannel is determined by measuring the SNR of each subchannel during startup. As the system is half-duplex there is no feedback signal back to the transmitter. Therefore a fixed number of bits is used instead of varying bit numbers.

The discrete complex multicarrier modulation signal [16] can be represented for n^{th} sample by:

$$s(nT) = \frac{1}{N}\sum_{k=0}^{N-1} A_k e^{j(2n\pi f_k T + \phi_k)} \tag{3}$$

The sampling frequency is $1/T$ and the period of one data symbol is $2NT$ in the inter-modulation frequency domain. The frequency, f_k, is given by:

$$f_k = f_0 + k(\Delta f) \tag{4}$$

where f_0 is the lowest frequency of the signal spectrum and $\Delta f = 1/NT$ is the frequency separation between carriers.

III. CHANNEL EQUALISATION

High bit rate, underwater acoustic, data transmission multicarrier modulation communication systems face high noise, strong reverberation and narrow bandwidth as well as multipath fading. There are two major effects of multipath fading: (i) magnitude decline; the received signal strength varies; (ii) intersymbol interference (ISI); the multipaths causes extension and distribution in the time domain. This causes intersymbol interference and is the major barrier to a high-bit rate data transmission system.

An adaptive equaliser updates its parameters in each period during the data transmission. The adaptive equaliser must track time variants in the channel response and adapt its coefficients to reduce ISI. The impulse response of underwater acoustic channel [17] is

$$h(\tau;t) = \sum_{k=1}^{L} a_k(t)\delta(\tau - \tau_k) \tag{5}$$

where $a_k(t)$ is L multipath factors, τ_t is the delay time. From (5), we can build a multipath model for an underwater acoustic channel. Then we use this model to analyse the effect of bit error with multipath fading.

The channel impulse response is given in (5). As mentioned by Chow et al. [17], the channel time-variant transfer function is $H(\omega;t)$. The channel response can be compensated by a transfer function $W(\omega;t)$, where it satisfies the equation $H(\omega;t)W(\omega;t) = B \approx 1$, where B is the DFT of $b(t)$ which has only $v+1$ contiguous non-zero samples. We assume the channel is non-time-varying during a short period, where multipath fading is the main problem.

Let $H(\omega;t) = H(\omega;t_0)$ and $H(\omega;t_0)W(\omega;t_0) \approx 1$, therefore $H(\omega)W(\omega) \approx 1$. Then, if we use an equaliser W with an adaptive algorithm (for example Least-Mean-Square or Recursive-Least-Square algorithm) to match the requirement $H(\omega)W(\omega) \approx 1$, the equalised channel becomes the ideal channel.

Fig. 1 shows the effect of multipath fading due to two additional paths with time delays $\tau_1 = 2\Delta T$ and $\tau_2 = 6\Delta T$ respectively, where Δ is the proportional delay time. The frequencies of direct path and two side paths are 50kHz. We assume no attenuation on the direct path while path 1 and path 2 have same attenuation of -6dB. It can be shown that the signal strength can vary by -15dB or more. The magnitude of the received signal is a maximum for zero delay of the two side paths because they are in phase with the direct path signal. The minimum point of the received signal occurs when the direct path signal is out of phase with each of the two side paths.

The latter choice is preferable since there are processors in the market which are empowered with special instructions for finite field arithmetic operations. Such instruction sets reduce the computational complexity of the decoding process. This is the error coding to be implemented on this modem.

IV. IMPLEMENTATION

A. Configuration

Fig. 5 shows a block diagram of the system. In the transmitter, a serial-to-parallel (s/p) buffer, adaptive threshold packets, MFSK modulator, LFM signal packet and parallel-to-serial (p/s) buffer are generated by a DSP. Also, synchronisation, s/p and p/s buffers, MFSK demodulator and threshold learning are performed by a DSP unit in the receiver.

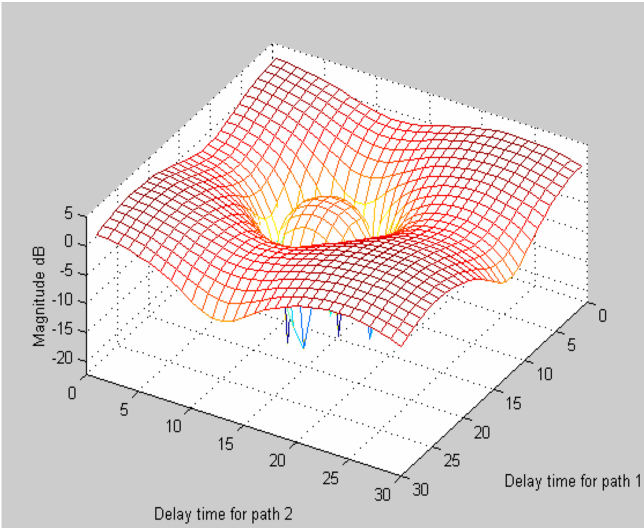

Fig. 1: The effect on the magnitude of the received signal due to the channel fading against the time delay of two paths

In the transmitter, the input data is via an RS232 link. The data is then converted from serial to parallel form by a serial-to-parallel buffer. The binary input data are parsed to each subchannel with one bit. Therefore there are 48 parallel bits represented by 48 frequency components, so the parallel data bits can be modulated by multicarrier modulation.

To overcome the multipath fading, eight adaptive threshold packets are added at the front of input data sequence before transmission. Each packet contains 48 bits. The bit patterns of adaptive packets are shown in Table 2. This is used as a reference signal for the receiver to estimate the channel, and then the IFFT algorithm is used for modulation.

TABLE 2
BIT SEQUENCE OF 8 ADAPTIVE THRESHOLD PACKETS

	Bit 1	Bit 2	Bit 3	Bit 4	……	Bit 45	Bit 46	Bit 47	Bit 48
Packet 1	1	0	1	0	……	1	0	1	0
Packet 2	0	1	0	1	……	0	1	0	1
Packet 3	1	0	1	0	……	1	0	1	0
Packet 4	0	1	0	1	……	0	1	0	1
Packet 5	1	0	1	0	……	1	0	1	0
Packet 6	0	1	0	1	……	0	1	0	1
Packet 7	1	0	1	0	……	1	0	1	0
Packet 8	0	1	0	1	……	0	1	0	1

As mentioned above, there are 48 frequency components within 43kHz to 53kHz. They are equally distributed in the frequency range, so that the frequency separation between carriers is

$$\Delta f = \frac{53k - 43k}{48} \approx 200Hz .$$

The choice of number of carriers is determined by the implementation in the DSP. The DSP uses 16 bit (one word) fixed point, so that multiples of 16 bits are used due to easy implementation. In the system, 48 bits (3 words) are used for each data packet. Also, the choice of the number of tones is a trade-off between the system sensitivity to multipath and practical implementation constraints [8].

After IFFT modulation, the data is converted from the frequency to time domain. At that time, the LFM signal is implemented at the front time slot of the data sequence. This is used for synchronisation in receiver.

In the receiver, some of the noise is reduced using a bandpass filter of 40kHz to 60kHz. The matched-filter captures the LFM signal from the received signal, which indicates the start of the data sequence. The output of this correlation provides an estimation of the channel that is used to select the pulse with the most energy for synchronisation. Then the data sequence is demodulated by the FFT algorithm. The data packets are converted from the time-domain to frequency-domain so that decoding can be take place in the error correction algorithm.

Once synchronisation has been achieved, threshold learning can be used to calculate the channel characteristic. Because the acoustic channel is a time varying channel, a fixed threshold cannot work well in the system. Therefore, the threshold of the detector is changed by transmitted reference packets at each data sequence. These change every 4.15s of the data sequence. Adaptive threshold in the frequency-domain is used and the training signal is sent out repeatedly after each time the LFM signal was send. At the receiver, the frequency-domain's threshold is calculated from the known training signal and the received training signal. According to the received training signal's frequency-domain characteristic, the adaptive threshold is calculated. This can be used to estimate the channel characteristic. Depending on the acoustic channel, an equaliser/echo canceller may be inserted into system at this point; otherwise the output of the FFT is passed to an error-correction algorithm [8].

B. Equalisation

In this implementation, a LMS algorithm is chosen. The channel equalisation is achieved by first identifying the short term channel characteristic and then an adaptive filter is used to extract the transmitted data. To identify the channel, M set of training sequences which consists of M chosen symbols is sent repeatedly over the channel. These M received sequences are averaged in order to eliminate received noise ξ. In the process the transmitter sends the known training sequence $x(t)$. After the transmission, the receiver receives the signal sequence. Here X and Y are the frequency domain representations of $x(t)$ and $y(t)$ respectively. Fig. 2 is the flowchart of a frequency domain version of equaliser (based on N sample DFT) [17]. The received signal in this equaliser is

$$Y = HX + \xi \tag{6}$$

where H is the channel characteristic and ξ is the received noise. After taking consecutive averaging, ξ can be neglected. Let W_k be the finite-impulse-response (FIR) equalizer at k^{th} estimation, we have

$$W_k Y = W_k HX = B_k X \tag{7}$$

where B_k is k^{th} estimation of the DFT of $b(t)$ which has only $v+1$ contiguous non-zero samples. The error of estimation can be defined by

$$B_k X - W_k Y = E_k$$

Then the design is to find the best W_k so that $\|E\|_2$ is minimised. Here a frequency domain LMS updating algorithm is used [17]. The LMS updating are:

$$B_{k+1} = B_k + 2\mu_b EX^* \tag{8}$$

and

$$W_{k+1} = W_k + 2\mu_w EY^* \tag{9}$$

where μ_b and μ_w are gain constant used to control the adaptive speed and the stability for the LMS algorithm.

The LMS iterative algorithm is actuated by the error signal E_k with an initial value of W being $W_0 = 1$. After the simulation, Fig. 3 shows the receiver decision error rate against the iteration number under different signal-to-noise ratios. The receiver decision error rate is reduced from 10^{-1} to $10^{-2} \sim 10^{-3}$ when using this channel equalisation algorithm.

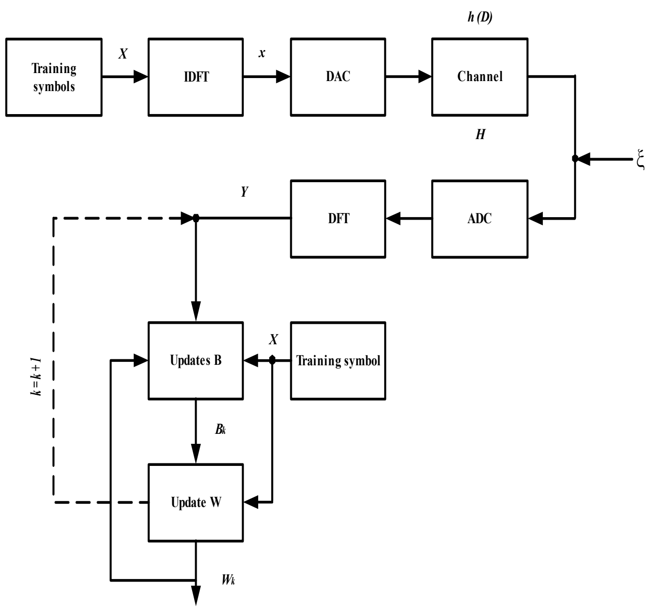

Fig. 2: Algorithm of the channel equaliser

C. Data sequence

As mentioned above, the data sequence is in packet form. The sequence contains synchronisation packets, a gap packet, adaptive threshold packets and information data packets. These packets are used to allow synchronisation and noise reduction. Each packet is formed by 48 frequencies within 43kHz to 53kHz, which represent 48 bits with a duration of 5.12ms.

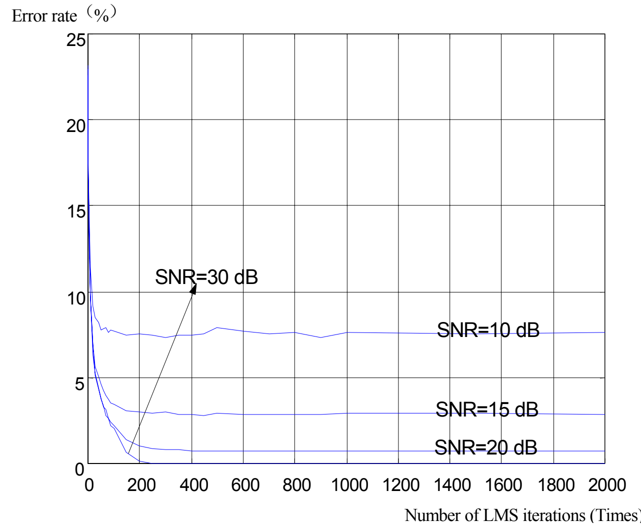

Fig. 3: The receiver decision error rate against with the iteration number under different signal-to-noise ratios using channel equalisation

The sequence begins with a Linear Frequency Modulation (LFM) signal packet which is used to synchronise the receiver to the start of the data. The details of LFM signal will be discussed in the next section. Then a packet for a gap signal follows the LFM packet so that synchronisation can be performed in this period. Eight adaptive threshold packets (ATP) are transmitted for receiver-training purposes. The training packets act as a reference of the transmitted signal block so that channel estimation can be calculated from the reference packets. Also, the long training sequence can be sufficient for system convergence. 800 information data packets (IDP) follow the adaptive packets. At the end of the data sequence, is another gap signal packet which can minimise the effect of multipath fading between two signal blocks. The time duration for one signal block is (1+1+8+800+1)*5.12ms = 4.15s. Fig. 4 shows the data sequence.

D. Synchronisation

For many systems, a Linear Frequency Modulation (LFM) signal is used for frame synchronisation. The most significant property of a linear frequency modulation signal is its symmetry in time and frequency. In general, the expression for a linear frequency modulation signal also referred to as a 'chirp' as mentioned by Rihazek [11] is:

$$s(t) = \cos(2\pi f_0 t + \pi k t^2) \tag{10}$$

The instantaneous frequency can be obtained by differentiation

$$f(t) = \frac{1}{2\pi} \cdot \frac{d}{dt}(2\pi f_0 t + \pi k t^2) = f_0 + kt \tag{11}$$

where f_0 is initial frequency, $|k| = \frac{B_w}{T}$, B_w is the bandwidth and T is the signal duration. Hence the LFM chirp described by Rihazek [11] is characterised by its starting frequency (f_0), stopping frequency (f_1), and time duration (T) as:

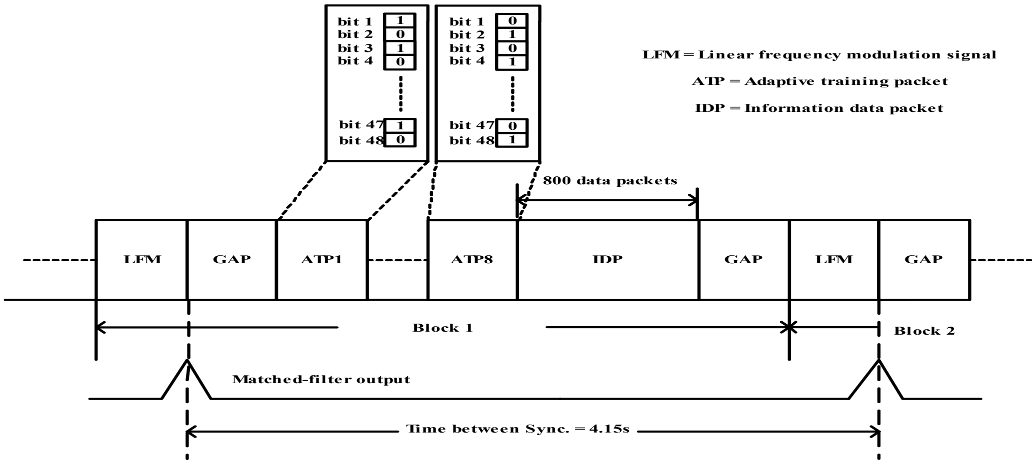

Fig. 4. Data sequence for one signal block

$$|k| = \frac{|f_1 - f_0|}{T} = \frac{B_w}{T} \tag{12}$$

The resolution of the time depends on the B/T factor.

At the receiver, a matched-filter is used to indicate the arrival of the LFM chirp. The output of this correlation allows selection of the channel with the most energy for synchronisation. The impulse-like auto-correlation function of the LFM signal allows synchronisation to be achieved by linear cross-correlation between the received signal and a known LFM signal. Therefore, the starting position of the data sequence can be found by the output of the matched-filter.

V. MULTIPATH FADING PROBLEM

In a shallow water channel, the multipath problem can affect the data rate and distance of communication. Although adaptive channel equalisation filters can be added to reduce the problem, when the carrier signal has fallen below certain noise floor, no equalisation can help. It is possible to reduce the channel capacity, which is equivalent to increasing the signal power spectrum. Another effective way is to assign multiple carriers to represent each bit in a symbol, and this will certainly decrease the channel capacity. For instance, let the i^{th} bit be sent with two carriers, f_1^i and f_2^i, and then recombine the signals at the receiving end.

The fading effect will then be averaged out. For instance, choose 8 carriers of magnitude 50, and $w(\tau)$ is the window function of length τ ; where $f^i = f_0 + i \cdot \Delta f$, $i = 1, \cdots, 8$; $f_0 = 43kHz$ and $\Delta f = 1.25kHz$. Then, the MFM signal becomes:

$$s(t) = \sum_{i=1}^{8} S_i \left(\sin \omega_i t\right) \cdot w(\tau) \tag{13}$$

The test channel has two multipaths with delay time $2\Delta T$ and $5\Delta T$, and with attenuation -6db each. S_i is the bit state of a symbol $\Omega = \{S_i\}$. Fig. 6 shows the fading of this test channel. The strength of carrier f_8 is reduced to about 20. If the channel capacity is now halved by allocating two carriers to a bit, then worst case fading is reduced to 25 (see Fig. 7). After the adaptive equalisation filter, it is improved significantly to 45 as shown in Fig. 8.

In order to reclaim the available channel capacity, the MFM technique can be improved by introducing a wavelet type MFM (WMFM). Now the S_i bit of a symbol Ω is represented by a set of carriers; $\{ f_1^i, f_2^i, f_3^i, .. f_m^i \}$. The WMFM becomes:

$$s(t) = w(\tau) \cdot \sum_{i=1}^{n} S_i \sum_{k=1}^{m} A_k^i e^{j(2\pi f_k^i t + \phi_k)} \tag{14}$$

Let $\psi(i,t) = w(\tau) \cdot \sum_{k=1}^{m} A_k^i e^{j(2\pi f_k^i t + \phi_k)}$ (15)

Therefore an information bit is transmitted over a number of carriers and hence the fading effect can be minimised. However, some inter-bit-carrier interference exists and that can affect the detection. The assignment of carriers to a symbol is an important issue. Let a symbol be $\Omega = \{0,1,1,1,0,1,0,1\}$ and let the bit assignment be $S_j \to \{f_j^j, f_{j+2}^j, f_{j+7}^j\}$, where the indices are computed with modulo 8. Let us choose $m = 3$; and $A_i^i = 1$ $A_2^i = 0.25$ and $A_3^i = 0.25$. Fig. 9 shows the relative signal strength. The minimum strength for those frequencies represent '1' is about 32, which is much higher than the previous example. Fig. 10 is the equalised result. If we set the threshold to 25, the original symbol can be recovered.

Alternatively, with this WMFM technique, the symbol can be recovered by the wavelet transform:

$$S_p^i = \int_{-\infty}^{+\infty} s(t) \psi(i,t,p) dt \tag{16}$$

In this case we set the translation $p = 0$. The discrete version of (16) is

$$S^i = S_0^i = \sum_{n=0}^{N-1} s(n) \psi(i,n,0) \tag{17}$$

The result is shown in Fig. 11. It is arranged so that ‘1’ and ‘0’ are mapped into the positive and negative indices with the magnitude representing the confidence level. Comparing Fig. 6 to 10, the result is far superior and the ‘1’ and ‘0’ are clearly detected.

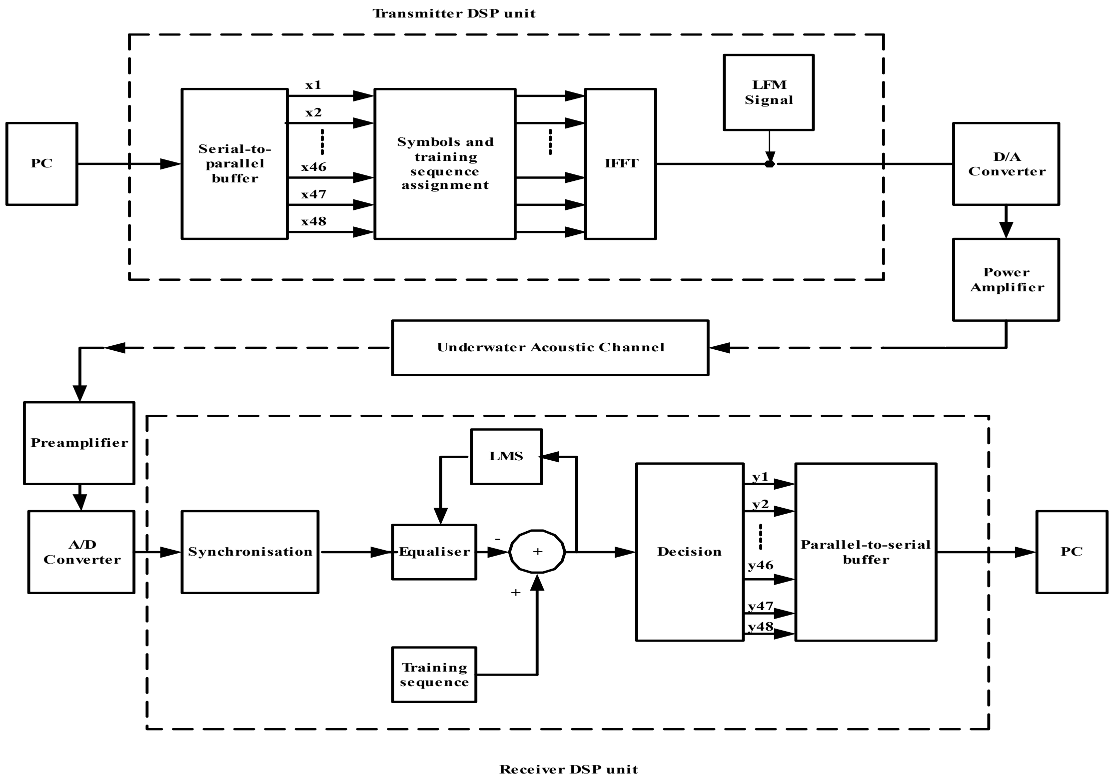

Fig. 5 System configuration

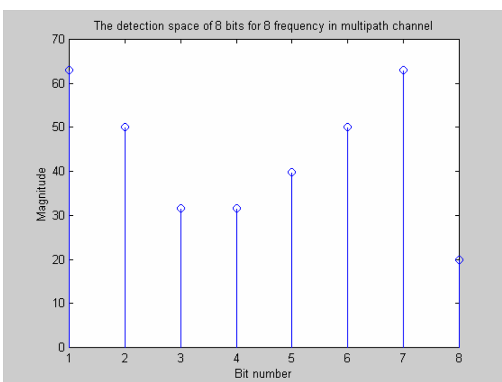

Fig. 6 Detection space of 8 bits for 8 frequencies in test channel

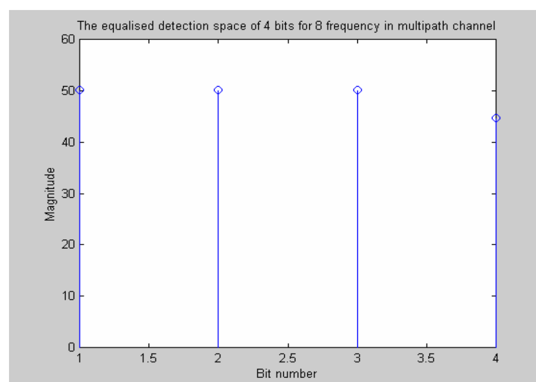

Fig. 8 Detection space of equalised 4 bits for 8 frequencies in test channel (filtered)

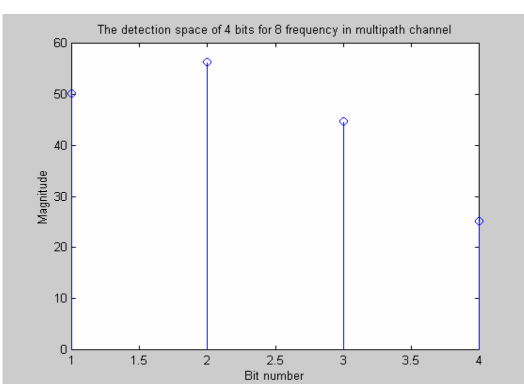

Fig. 7 Detection space of 4 bits for 8 frequencies in test channel

Fig. 9 Detection space of each bit into 3 frequencies in test channel

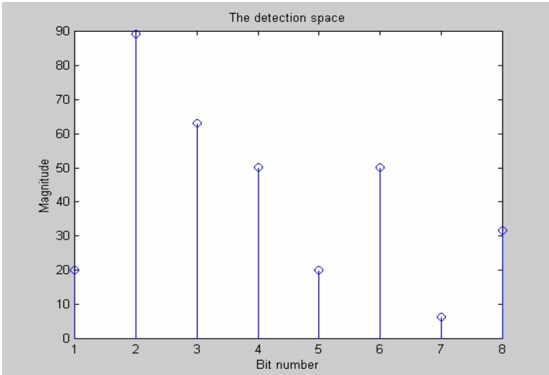

Fig. 10 : Equalised Detection space of each bit into 3 frequencies in test channel(filtered)

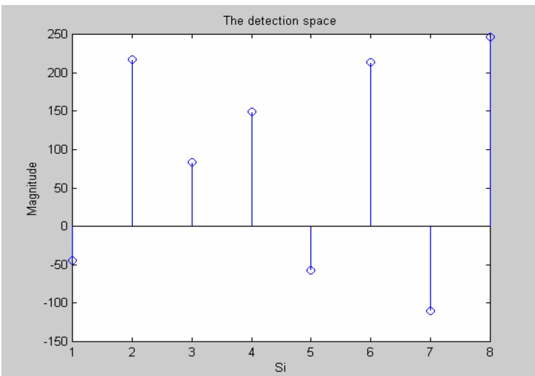

Fig.11 Detection space of each bits using 3 wavelets

VI. RESULTS

The system has been tested in CityU's swimming pool and in coastal areas near Hong Kong. The pool has dimensions of 50m x 25m, and the transducers are at a depth of 0.6m. There is serious multipath fading problem, and the reflected signals from the side walls and the bottom of the pool are strong. A number of open water sea trials have also been carried out. One took place at Da Mei Do, in the New Territories of Hong Kong (Fig. 12). The distance between the transducers varied between 600m and 1.2km, and the water depth between 3.5 and 8m. The measured data rate was 10kbps.

In one test, the transducers were at a depth of 4 metres. The receiver was at the wharf and the transmitter on the boat. The depth of the water was 5.6 metres at the wharf and 8 metres at the boat. The distance between two transducers was 820m, measured by GPS. Fig. 13 shows the results of sending an uncompressed bit-mapped picture of 24 bits 80*60 pixels. The top left hand picture is the sent signal. The others are from the received signal. The error rate of the raw data was less than 5%. In these trials there was no error correction, channel equalisation, or system identification being used.

Current work is concentrating on implementation of the equalisation, identification and wavelet detection algorithms to be programmed into the DSP, along with the error correction routines. It is expected that the effective data rate will be around 8kbps with a BER of 1 in 10^4, thus allowing compressed video to be transmitted.

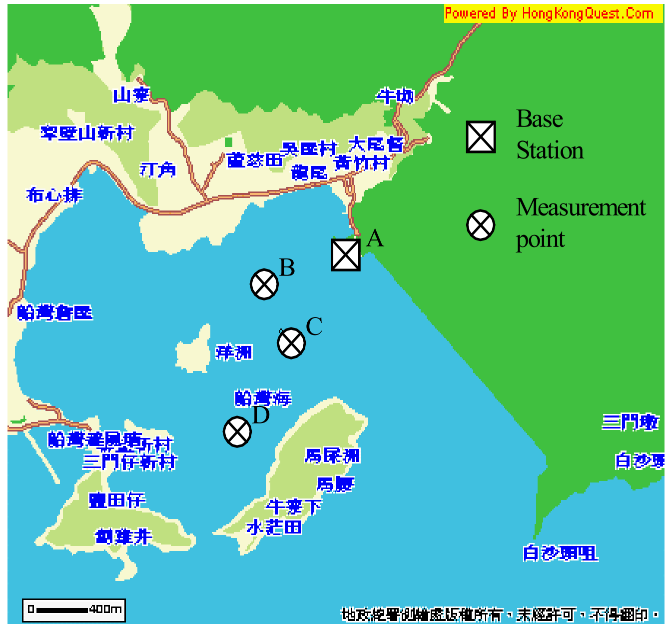

Fig. 12. The map of Da Mei Do

Fig. 13. Received pictures from sea trial

VII. CONCLUSIONS

A modem designed for underwater use has been described. This uses multi-carrier modulation over 48 frequencies allowing for use in multipath environments, such as shallow water. It has been tested at a data rate of 10kbps over a range of 1km, and without implementation of error correction or equalisation techniques, has a bit error rate of 5%. A new wavelet detection scheme is developed with promising results. Experimental results in the laboratory, as well as simulation, show that this BER can be reduced by a factor of 10^3 when fully developed.

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